

Critical-mode control stabilizes switch-mode power supplies

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A switch-mode power supply (SMPS) can operate in two conduction modes, each one depicting the level of the current circulating in the power choke when the power switch turns on. Two "black-box" supplies, which deliver the same power levels but work in different conduction modes, have properties that differ dramatically during dc and ac conditions. The conduction mode also affects the amount of stress on the power elements. Most low-power flyback SMPSs, such as those in off-line cellular battery chargers and VCRs, operate in the discontinuous area. Maintaining operation in this area improves the supply's ability to respond to changes in the load and makes it easier to design the power supply's compensation network.

Figures 1a and b show the general shape of the current flowing through the converter's coil during a few representative cycles. Current ramps up when the switch closes. During this on time, a magnetic field builds up in the inductor's core. When the switch opens, which is the off time, the magnetic field collapses, and, according to Lenz's law, the voltage across the inductance reverses. In this case, the current has to find some way to continue its flow and begin its decrease—for example, through the output network of a flyback converter or through the freewheel diode in a buck converter.

If the switch turns on again during the ramp-down cycle and before the current reaches zero (Figure 1a), the supply is operating in continuous-conduction mode (CCM). Alternatively, if the energy-storage capability of the coil is such that its current dries out to zero during the switch's off time, the supply is operating in discontinuous-conduction mode (DCM, Figure 1b). The amount of dead time for which the current stays at a null level defines how strongly the supply operates in DCM. If the current through the coil reaches zero and the switch turns on immediately (no dead time), the

Discontinuous-conduction-mode operation eases switch-mode power-supply design and allows for better response to step-load changes. Operating at the critical-conduction point—beginning a new cycle at the exact point the inductor current falls to zero—ensures that the converter stays in discontinuous-conduction mode.

converter operates in critical-conduction mode.

Where is the boundary?

You can think of the boundary between the modes and the three corresponding design approaches in three ways. The first design approach is to define the critical value of the inductance, L_C , for which the supply works in either CCM or DCM, given a fixed nominal load. The second approach

deals with a known inductance, L , and requires answers to the following questions: What level of load, R_C , pushes the supply into CCM? Or, what should the minimum SMPS load be before entering DCM? The third approach uses fixed values of L and R but adjusts the operating frequency, F_C , to stay in critical conduction.

A few lines of algebra that correspond to a standard flyback converter help derive the three equations for L_C , R_C , and F_C (Figure 2). Keep in mind some key points about this converter:

- The average inductor voltage per cycle should be null;
- from Figure 1b, when $L=L_C$, $I_{L(AVG)}=2 \times I_P$;
- and 100% efficiency leads to $P_{IN}=P_{OUT}$.

You can determine the dc-voltage transfer ratio in CCM using the first point above and equating the areas in Figure 2b: $V_{DD} \times N \times D = V_O \times (1-D)$, where D is the duty cycle and N is the primary/secondary turns ratio. After factoring,

$$\frac{V_{OUT}}{V_{IN}} = \frac{D}{1-D} \times N. \quad (1)$$

As you can see from Figure 1b, the flux that the coil stores during on time is down to zero at the beginning of the next cycle, when the inductance equals its critical value ($L=L_C$). You can mathematically express this statement by integrating the formula as follows:

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$$V_L \times dt = L \times dI_L$$

Thus,

$$\int_0^D \frac{D}{F_{SW}} V_L \times dt = L_C \times \int_0^{I_P} dI_L$$

From point 2,

$$\frac{V_{IN} \times D}{F_{SW}} = L_C \times I_P = 2 \times I_{L(AVG)} \times L_C$$

From point 3,

$$V_{IN} \times I_{L(AVG)} = I_O \times (V_{IN} \times N + V_{OUT}),$$

or

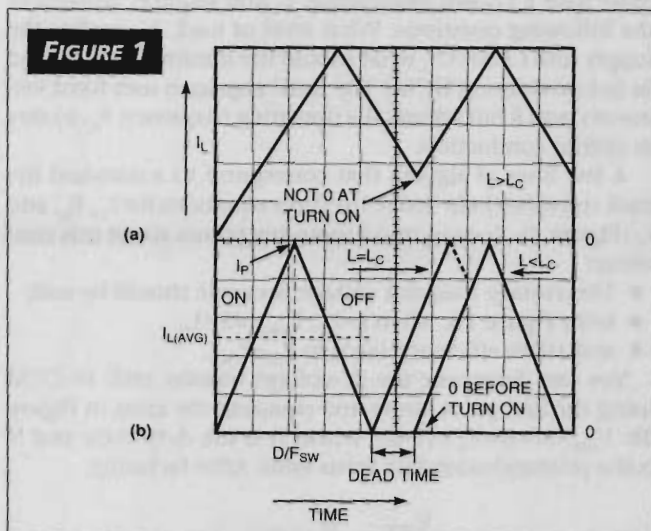
$$I_{L(AVG)} = I_O \times \left(N + \frac{V_{OUT}}{V_{IN}} \right)$$

By definition, $I_O = V_{OUT}/R$, and from Equation 1, $V_{OUT} = V_{IN} \times N \times D / (1-D)$. If you introduce these elements in the above equations, you can solve for the critical values of L_C , R_C , and F_C as follows:

$$L_C = \frac{R_C \times (1-D)^2}{2 \times F_{SW} \times N^2};$$

$$R_C = \frac{2 \times L_C \times F_{SW} \times N^2}{(1-D)^2};$$

$$F_C = \frac{R_C \times (1-D)^2}{2 \times L_C \times N^2}.$$

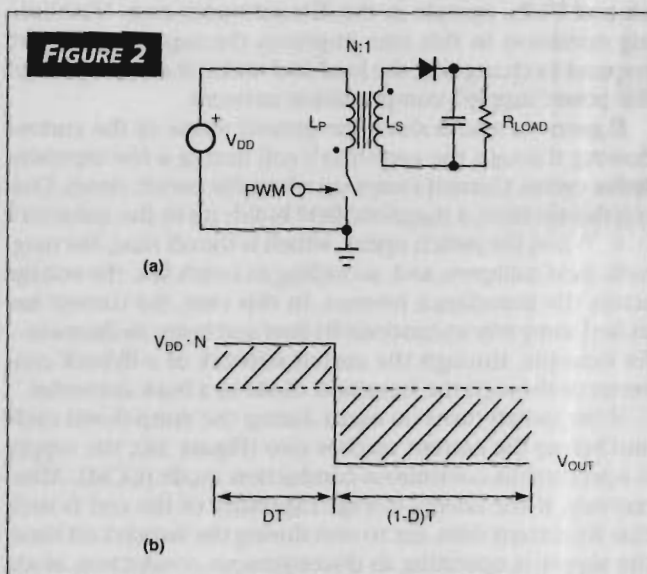


For a supply operating in the continuous-conduction mode, a new cycle starts before the current drops to zero (a). In discontinuous conduction (b), the current drops to zero, and some dead time may exist before the next cycle. If the switch turns on immediately when the current reaches zero, the supply is operating in critical-conduction mode.

The flyback converter, as with the boost and buck-boost structures, has an operating mode comparable to someone filling a bucket (coil) with water and then flushing it into a water tank (capacitor). In this analogy, the user first presents the bucket to the spring (on time) until its inner level reaches a defined limit. Then, the user removes the bucket from the spring (off time) and flushes the water into a tank that supplies a fire engine (load). The bucket can be totally empty before refilling (DCM), or some water can remain before the user presents the bucket back to the spring (CCM). Suppose that the user is experienced, thus ensuring that the recurrence period (on plus off time) is constant.

The end user is a firefighter who closes the feedback loop via his voice, shouting for more or less flow to the tank. If the flames suddenly increase, the firefighter requires more power from the engine and thus asks the bucket operator to provide the tank with a higher flow. In other words, the bucket operator fills his container longer (on time increases). However, because from experience this user keeps his working period constant, the time he spends flushing water into the tank naturally diminishes (off time decreases), as does the amount of water poured. The fire engine runs out of power, making the firefighter shout louder for more water, extending the filling time, and so on. Thus, the loop oscillates.

This behavior, which severely affects overall dynamic performance, is typical for converters in which the energy transfer is not direct (unlike the buck-derived families). In the time domain, a large step-load increase requires a corresponding percentage rise of the inductor current. This rise necessitates a temporary duty-cycle augmentation, which, with only two operational states, causes the diode conduction time to diminish. Therefore, this load increase implies an initial decrease in the average diode current, rather than an increase, as desired. When the converter is operating heavily,



For a basic flyback converter (a), the voltage across the secondary coil (b) depends on V_{DD} , the primary/secondary turns ratio, N , and the duty cycle, D .

SPICE CONVERTER SIMULATIONS

You can distinguish between two families of converter Spice models: average and switching. Average models implement either the state-space-averaging technique or Vorperian's solution (Reference 4). These average models require a short computational time and can work in ac or transient analysis because they use no switching components. Some of these models support large transient sweeps; others accept only small-signal conditions.

Alternatively, switching models are the Spice reproduction of a breadboard, and you use these models to simulate the supply using your selected PWM controller or the MOSFET model from its manufacturer.

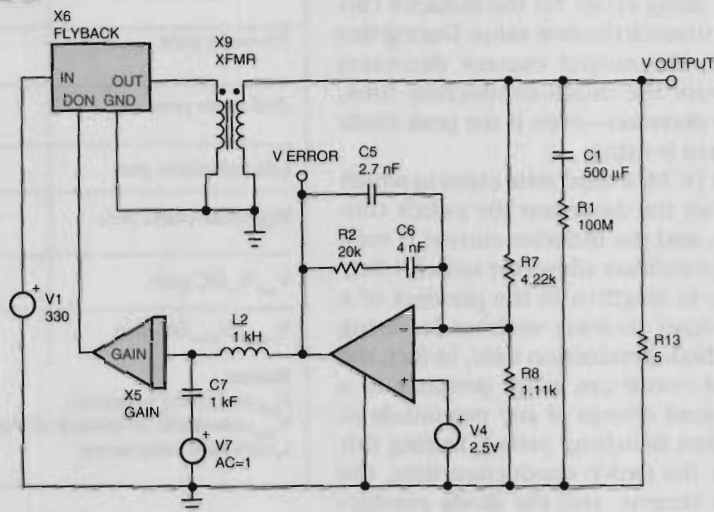
Both models have advantages. Average models simulate quickly but include no leakage-energy spikes or parasitic-noise effects. Switching models take longer to run because the simulator has to perform a thin analysis (internal step reduction) during each commutation cycle. However, because switching models can include parasitic elements, these models allow you to dive into the nitty-gritty of the converter under study.

Spice models are available from several sources, and you can also write your own (Reference A). Intusoft (<http://205.147.9.162/>) recently released an SIMPS library, which includes numerous average and switching models. Among these models, a model originally developed by Sam-Ben-Yakov from Ben-Gurion University (Israel) converges well and finds its own dc operating point. This model also allows ac simulations as well as large-signal sweeps (Listing A).

Implementing the model in Figure A is straightforward; this circuit model conforms to Ridley Engineering's (<http://members.aol.com/ridleyeng/index.html>) Power 4-5-6, 100-kHz SMPS. The circuit is perfectly suited to model the converter operating in either discontinuous or continuous mode. The model automatically switches between both modes. You can temporarily open the loop and conduct ac simulations to isolate the error amplifier and adjust the compensation network until the circuit meets the specifications.

The fastest way to open the

FIGURE A



The model of a 100-kHz flyback SMPS includes an error-amplifier model, a block for PWM gain (X5), and a flyback-converter IC model (X6).

loop is to include an LC network, such as L2-C7. The inductor maintains the dc error level such that the output stays at the required value, but the inductor stops any ac error signal that would close the loop. The capacitor provides for ac signal injection, thus allowing a normal ac sweep. For ac sweeps, let L2=1 kH and C7=1 kF. Now, if you want to run a transient analysis, decrease L2 to 1 nH and C7 to 1 pF. This method provides the advantage of an automatic dc duty-cycle adjustment and allows you to quickly modify the output parameter without tweaking the duty source at every change.

You can derive the error-amplifier model directly from the specifications on the selected controller's data sheets. You can also build a simplified macro-model or a more detailed component model.

The X5 subcircuit simulates the gain of the PWM. This block converts a dc voltage (the error-amplifier voltage) into a duty cycle, D. Average models accept as much as 1V as a duty-cycle control voltage (D=100%). However, the actual IC's sawtooth oscillator generally can swing as high as 3 or 4V, which forces the internal PWM stage to deliver the maximum duty cycle when the error amplifier reaches this value. Thus, to account for the 1V-maximum input of the average models, the insertion of an attenuator with a $1/V_{SAW}$ ratio after the error-amplifier output is mandatory. For example, if the sawtooth amplitude of the IC you use is 2.5 V_{pp}, then the ratio is $1/2.5=0.4$.

LISTING A—FLYBACK-CONVERTER ISSPICE4 NETLIST

```
.SUBCKT FLYBACK DON IN OUT GND {FS=??? L=??? N=???}
BGIN IN GND I=(VLM)*V(DON)/(V(DON)+V(DOFF))
BELM OUT1 GND V=(IN)*V(DON)-V(OUT)*V(DOFF)/(IN)
RM OUT1 5 1M
LM 5 8 {L}
VLM 8 GND
BGOUT GND OUT I=(VLM)*V(DOFF)/(IN)/(V(DON)+V(DOFF))
VCLP VC 0 9M
D2 VC DOFF DBREAK
D1 DOFF 6 DBREAK
R4 DOFF 7 10
BDOFFM 6 GND V=1-V(DON)-9M
BDOFF 7 GND V=2*(VLM)*(FS)*{L}/V(DON)/V(IN)-V(DON)
.MODEL DBREAK DT=1N CJO=10P N=0.01
.ENDS
```

Reference

A. Basso, C. "Write your own generic Spice power supplies controller models, parts I and II," PCIM, April/May 1997.

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ily in continuous mode and if the inductor-current rate is small compared with the current level, it may take many cycles for the inductor current to reach the new value. During this time, the output current decreases because the diode conduction time, T_{OFF} , decreases—even if the peak diode current is rising.

In DCM, a third state exists in which neither the diode nor the switch conduct, and the inductor current is null. This condition allows the switch's duty cycle to lengthen in the presence of a step-load increase without lowering the diode conduction time. In fact, the DCM circuit can adapt perfectly to a step-load change of any magnitude in the first switching period. During this time, the switch conduction time, the peak current, and the diode conduction time simultaneously increase to the values that the converter maintains indefinitely at the new load current.

Delay adds a right-half-plane zero

You can mathematically describe the extra delay inherent to the coil charge/discharge process as a right-half-plane zero (RHPZ) in the transfer function of the form

$$A_v = \frac{(1 - s_{Z1})x \dots}{\dots}$$

This delay forces the designer to roll-off the loop gain at a point at which the phase margin is still secure. Actually, a

TABLE 1—FLYBACK-CONVERTER POLE, ZERO, AND GAIN CALCULATIONS

	DCM	CCM
1st-order pole	$\frac{2}{2 \times \pi \times R_{LOAD} \times C_{OUT}}$	
2nd-order pole	High-frequency pole, see Reference 4	$\frac{(1-D)}{2 \times \pi \times \sqrt{L_p} \times C_{OUT}}$
Left-half-plane zero	$\frac{1}{2 \times \pi \times R_{ESR} \times C_{OUT}}$	$\frac{1}{2 \times \pi \times R_{ESR} \times C_{OUT}}$
Right-half-plane zero	High-frequency RHPZ, see Reference 4	$\frac{R_{LOAD} \times (1-D)^2}{2 \times \pi \times L_p \times D}$
V_{out}/V_{in} DC gain	$N \times D \times \sqrt{\frac{R_{LOAD}}{2 \times L_p \times F_{SW}}}$	$\frac{D}{(1-D)} \times N$
V_{output}/V_{error} DC gain	$\frac{V_{IN}}{V_{SAW}} \times \sqrt{\frac{R_{LOAD}}{2 \times L_p \times F_{SW}}}$	$\frac{V_{IN}}{V_{SAW}} \times \left(1 + \frac{V_{OUT}}{V_{IN}}\right)^2$

Notes:
 F_{SW} =switching frequency.
 V_{SAW} =sawtooth amplitude of the oscillator's ramp.
 L_p =primary inductance.

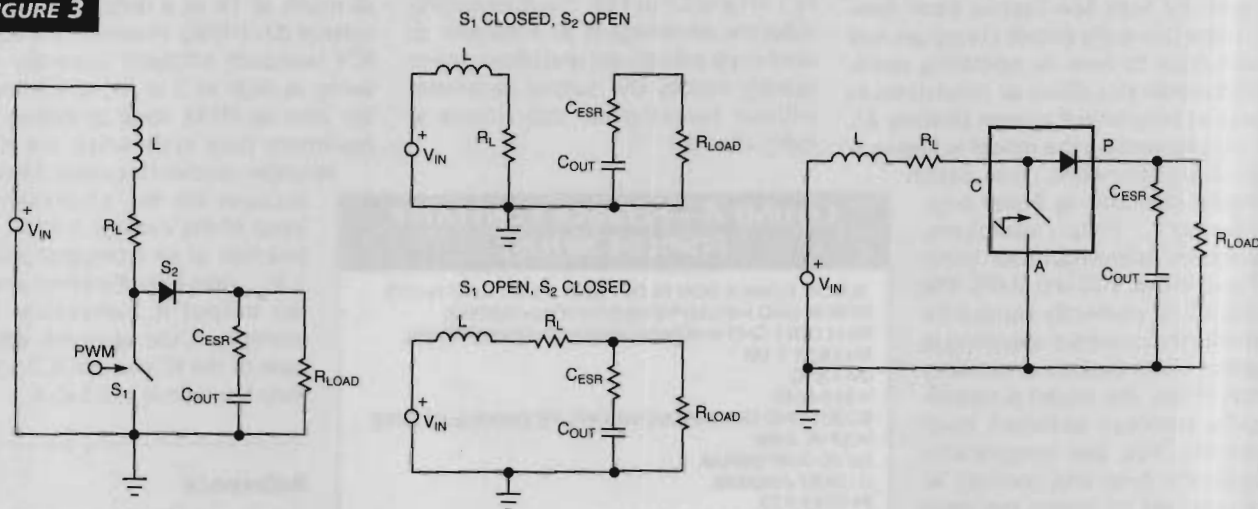
classical zero in the left half-plane has the form

$$A_v = \frac{(1 + s_{Z1})x \dots}{\dots}$$

and provides a boost in gain and phase at the insertion point. Unfortunately, the RHPZ in this SMPS case gives a boost in gain but lags behind the phase. More viciously, the RHPZ's position moves as a function of the load, which makes determining compensation an almost impossible exercise.

Rolling-off the gain well under the worst RHPZ position is the usual solution. The low-frequency RHPZ is present only

FIGURE 3



You can model an SMPS using a state-space-averaging approach (b) and PWM-switch-model approach (c). The state-space-averaging technique breaks down the switching action of the converter in (a) into two stable positions of the switches in (b) before analysis can take place.

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in flyback-type converters—such as boost and buck-boost—that operate in CCM. When the power supply enters DCM, the RHPZ moves to higher frequencies and thus becomes negligible, which eases loop compensation. For additional information, **Reference 1** offers an interesting experimental solution to cure the boost-topology converter of its low-frequency RHPZ.

Model a converter

You can see the effect of this RHPZ by modeling the converter. Two main modeling approaches study a converter's ac and dc characteristics. The well-known state-space-averaging (SSA) method leads to average models (**Reference 2**). In modeling, a set of equations describes the electrical characteristics of a switching system for the two stable switch positions, such as those of a boost-type converter (**Figures 3a and b**).

The SSA technique consists of smoothing the discontinuity associated with the transition between these two states, then deriving a model for which a unique state equation that describes the average behavior replaces the converter's switching component. The result is a set of continuous nonlinear equations in which the state coefficients now depend on duty cycles D and D' ($1-D$). A linearization process finally leads to a set of continuous linear equations. **Reference 3** contains an in-depth and pedagogical description of these methods.

As you can see from **Figure 3b**, the SSA technique models the converter in its entire electrical form. In other words, the process should carry over all of the converter elements,

including various in/out passive components. Depending on the converter structure, the process can be long and complicated.

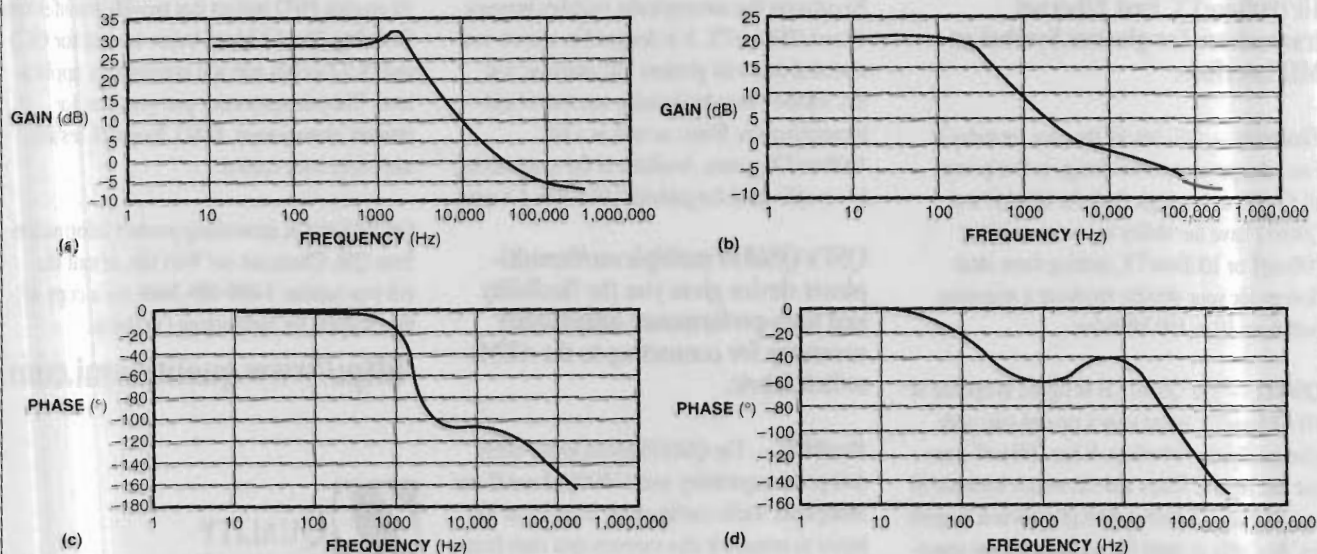
The second modeling approach uses a PWM switch model, independently developed by Vatché Vorperian of the Virginia Polytechnic Institute and State University (Blacksburg, VA, **Reference 4**) and Larry Meares of Intusoft (<http://205.147.9.162/>, **Reference 5**). Using this method, you simply model the power switch alone and then insert an equivalent model into the converter schematic—in exactly the same way as when studying a bipolar amplifier's transfer function (**Figure 3c**). Both authors of this method demonstrate that the flyback converter operating in DCM is still a second-order system affected by a second high-frequency pole and an RHPZ. **Reference 6** details how to run simulations using these models; for additional information, see box "Spice converter simulations."

Flyback-converter Bode plot

The authors of the above studies extracted the poles and zeros of converters operating in DCM and CCM, providing a designer with the necessary insight to make a power supply stable and reliable. **Table 1** summarizes the pole/zero positions as a function of the operating mode and also specifies the various gain definitions for a flyback converter.

You can generate the Bode plots using a multitude of manual methods or, in a more automated way, using dedicated software, such as Power 4-5-6 (Ridley Engineering, <http://members.aol.com/ridleyeng/index.html>). For example, you can ask this program to design two 100-kHz

FIGURE 4



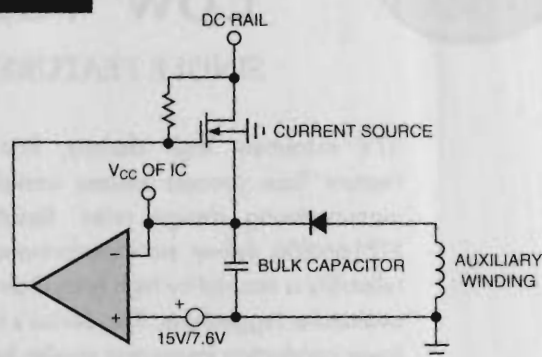
Gain and phase plots of the same converter operating in CCM, (a) and (c), and in DCM, (b) and (d), reveal that a DCM converter requires a simple two-pole, single-zero compensation network. However, a two-pole, two-zero network is necessary to stabilize a CCM converter.

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voltage-mode SMPs that operate in different modes but have equivalent output power levels. The resulting Bode plots (**Figure 4**) include the high-frequency pole and RHPZ in DCM, which **Reference 4** describes. The plots clearly indicate that the DCM converter requires a simple double-pole, single-zero compensation network (type-2 amplifier). However, a two-pole, two-zero, type-3 amplifier is mandatory to stabilize the CCM converter. Furthermore, the CCM's second-order pole moves in relationship to the duty cycle, whereas the poles and zeros in a DCM supply are fixed.

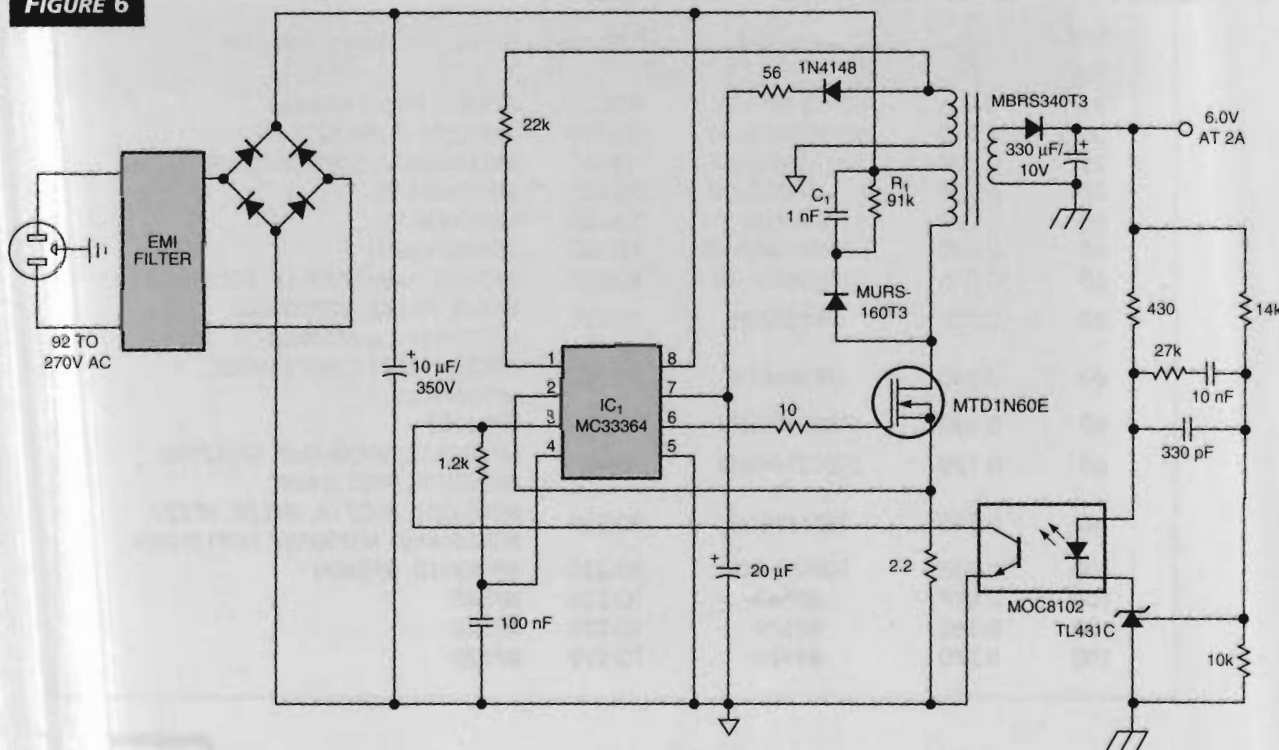
Stay in critical mode

Thus, keeping an SMPS in DCM eases compensation-network design. This operational mode also ensures stable and reliable behavior as long as the converter stays in the discontinuous area. Two options exist to ensure that the converter stays in DCM, regardless of the load span at the output. The first, for which you need to know all load conditions, is to calculate L_p such that the converter always stays in DCM. The second approach is to adjust the switching frequency to stay permanently in DCM, regardless of the load level. Motorola's (www.mot.com) MC33364 critical-mode controller uses the second approach. This critical-conduction controller ensures that a switch turns on immedi-

FIGURE 5

To eliminate a start-up resistor from the design, the MOSFET charges the bulk capacitor until the voltage on its pins reaches the start-up threshold. Then the MOSFET opens, and the circuit operates on its own.

ately after the primary current drops to zero. In this case, you do not have to worry about the values of the load, because the controller tunes its frequency to keep the SMPS

FIGURE 6

In a 12W ac/dc wall-adaptor supply, which operates in critical-conduction mode with the help of controller IC₁, R₁ and C₁ clip leakage-energy spikes and smooth the rising drain voltage, which limits the radiated noise.

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in DCM. Stability is then guaranteed over the full load range.

To maintain operation in critical-conduction mode, the controller needs to know the level of the primary current. The most economical option uses the signal from the auxiliary winding. When this signal falls to zero, the controller initiates a new cycle. As a safeguard against a loss of the synchronization signal, the controller restarts the converter if the driver's output stays off more than 400 μ sec after the inductor current reaches zero.

As stated, this type of critical-mode control involves adjusting the frequency to maintain operation in DCM. So, when using this type of critical-mode control, you may want to provide another safeguard to prevent the operational frequency from shifting to a high value, which can happen in the absence of a load. A frequency that is too high engenders unacceptable switching losses and makes the design of the EMI filter difficult. One version of the MC33364 includes an internal frequency clamp whose function is to limit the maximum excursion.

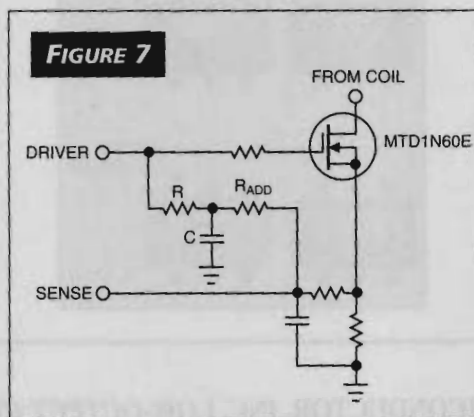
Good riddance, start-up resistor

Most off-line SMPSs self-start. A start-up resistor charges a bulk capacitor until reaching the switching IC's undervoltage limit. As the bulk-capacitor voltage begins to decrease, the circuit starts to actuate the switching transistor, and the auxiliary supply feeds the controller back through the rectifier. Once the circuit reaches the steady-state level, the start-up resistor is still in the circuit and wastes substantial energy in heat.

When designing a low-power SMPS, eliminating any source of wasted power raises overall efficiency. Figure 5 shows one method that the MC33364 uses to quash the start-up element. When you first apply power, the MOSFET charges the bulk capacitor until the voltage across the capacitor reaches the start-up threshold of 15V. At this time, the MOSFET opens, and the circuit operates by itself.

Figure 6 illustrates a 12W ac/dc wall adapter that uses the critical-conduction technique. The R_1 and C_1 network clips leakage-energy spikes and also smoothes rising drain voltage, correspondingly limiting the radiated noise. This last feature is unfortunately no longer valid when you use a clipping circuit comprising a fast rectifier and a zener diode. A leading-edge blanking network largely diminishes the circuit's sensitivity to the noise on the sense resistor. This network blanks the starting portion of the primary ramp-up current, which can be the seat of spurious spikes that stem from a resonance between the parasitic interwinding capacitors and the on gate-source current.

Because all current-mode converters are inherently unsta-



By integrating the driver's output, current-ramp compensation can stabilize the converter.

ble over a certain duty-cycle value, it's wise to add some current-ramp compensation, even in DCM (Reference 7). Even though no oscillator pinout is available on the MC33364, you can still provide the sense input with some ramp by integrating the driver's output (Figure 7). The resulting linear ramp adds to the sense information, which stabilizes the converter. You can also adopt this method in other cases and when using different controllers, even when the oscillator's ramp is available. The integrator option prevents external loading of the internal oscillator, which, in certain circumstances, can lead to erratic behavior.

EDN

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